# DESIGN AND IMPLEMENTATION OF A FIBRE OPTICS LINK

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By
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## Certificate

This is to certify that the present work entitled, "Design and Implementation of a Fibre Optics Link" has been done under my supervision and has not been submitted elsewhere for a degree.

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## ABSTRACT

A fibre optics link operating at 18 Mbps has been implemented. The present work consisted mainly of designing and fabricating the low noise receiver amplifier capable of working at that Bit Rate. Different circuit configurations were considered, and according to the design considerations as detailed in the report, a BJT preamplifier in transimpedence configuration was implemented as the front end amplifier. The complete receiver amplifier consisted of the preamplifier, a linear amplifier and the decision circuit. A bit synchromiser was also fabricated to operate at the particular bit rate for extracting the clock information from the data. Crystals were used at the transmitter and the receiver to improve the acquisition time of the link. In this report the performance of the complate receiver has also been discussed and analysed.

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# DESIGN AND IMPLEMENTATION OF A FIBRE OPTICS LINK

## Chapter 1 Introduction

### 1.1 Fibre Optic Communication

Optical fibres have opened up a new domain in communication systems engineering, especially at high data rates and electrically noisy environments. It is poised to take over as the main communication link for telephone trunks as the copper conductor systems are not able to cope up with the very high volume of traffic. Optical links are also becoming important medium of communication between computers - a number of networks have come into operation utilizing the excellent properties of the fibre. The telephone trunks made of optical fibres can be easily used to send voice and data simultaneously, as a result of its high capacity, and thus, Integrated Service Digital Networks can be realised. Video signals also can be accommodated in the link. There are active research worksgoing on all around the world to harness the capabilities of the fibres - new improvements in fibres, its sources and detectors and the interface electronics is announced almost every alternate day.

Optical fibres, as the name suggests, is a medium of communication where the information is transmitted in the form of light. Light acts as the carrier in this

transmission - which results in the inherent high bandwidth of the fibres. In most systems today, the utilization of the fibre bandwidth is limited by the electronic interface Data Rates upto several Gigabits have been reported - for example, an experiment in Japan has achieved 2 Gbps with the fibres [1].

The nature of transmission medium has resulted in another important characteristic of the fibre - it is not affected by Electromagnetic Interferences or Radio Frequency Interferences. The fact that light, though an electromagnetic wave in a generalised sense, is not affected by electric and magnetic fields as the electric signals are, makes fibres a far better communication medium than any copper conductor cables. This has made the fibre a very popular link in tactical systems, where an ultra reliable link is a necessity. Also, power stations, where a lot of noise is generated by switches and motors, prefer fibres to transmit signals. The transmission of optical signal cannot be monitored external to the fibre - as can be done with electrical signals in cables or wires - thus, optical link offers more security to the users.

Fibres also have important mechanical advantages over copper cables - they are lighter, smaller and more flexible. Therefore, fibre cables can be laid more easily. The small size of the fibres has resulted in its application in many areas as the monitoring devices - the areas range

from medical systems to nuclear test reactors.

The fibres are now being used for long haul systems like submarine cables, though traditionally it was more popular for shorter distance systems. The average repater span in the submarine systems is 30 km. Laboratory experiments have shown that the fibres can be used for much larger distances - a 134 km link has be successfully implemented in Japan [2] and an 161.5 km link at Bell Laboratories [3] . All these are the result of very low loss fibres, which is also decreasing day by day.

A disadvantage with the fibre is the couplers - the couplers introduce a good amount of loss in the system with low loss fibres. For this reason, the fibre links tend to be point to point instead of any other topology. In Local Computer Networks also, a number of new networks have come up and new protocols proposed for the same reason. This networks tend to use the fibre in a broadcast type of network while retaining the advantages of high bandwidth and low noise. Also, the fibre links tend to be unidirectional, which has also been taken care of a number of topologies.

## 1.2 A Typical Fibre Optics Link

A typical fibre optics link used in digital transmission systems is shown in Figure 1.1. The basic blocks in the transmitter and receiver modules have been

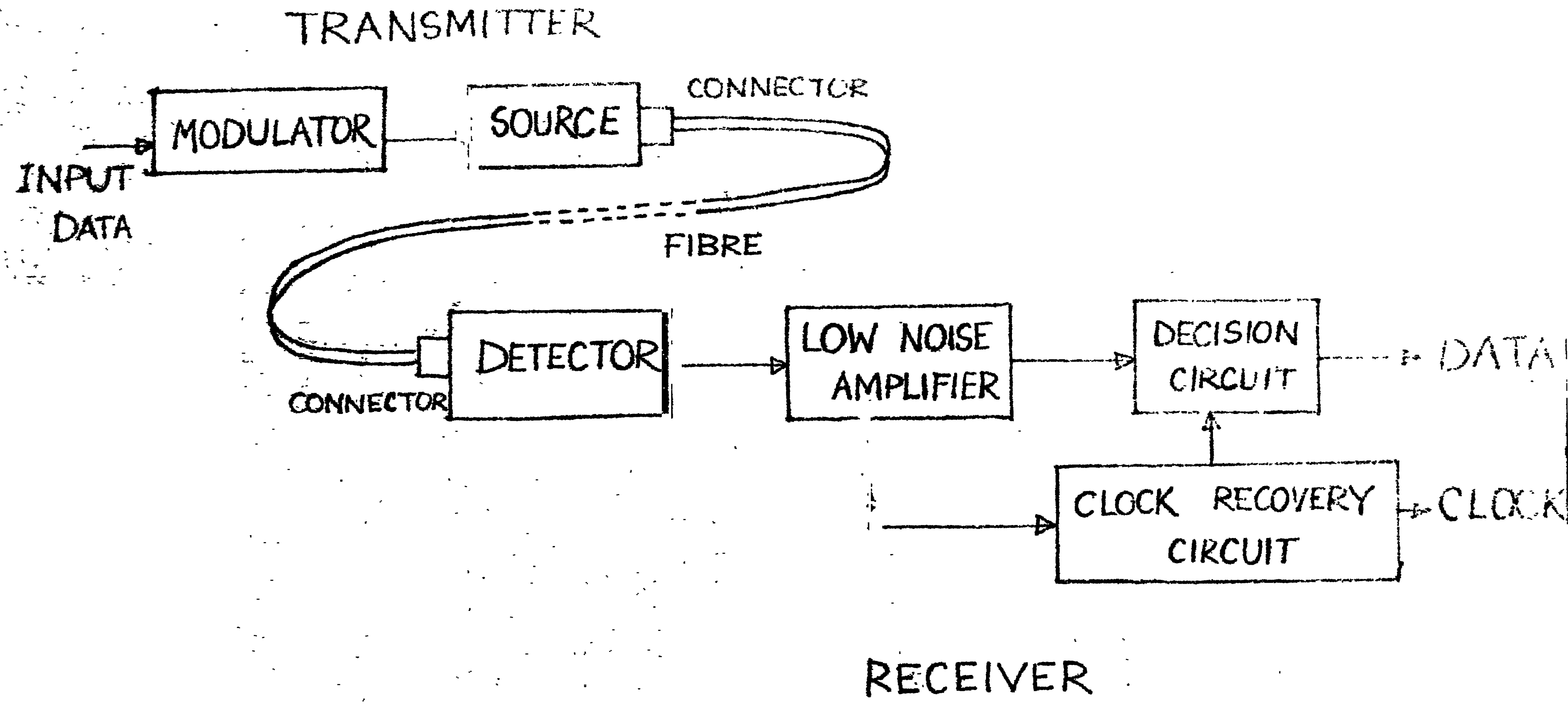



FIG 1.1. A FIBRE OPTIC LINK

shown. The transmitter consists of mainly two parts - the source is the source of optical signals in the link and the modulator is used to modulate the light source according to the incoming data.

In the receiver, a detector converts the optical signals into electrical signals and a following low noise amplifier amplifies the signal and shapes it. The clock, in digital transmission, is derived from the data itself to increase the efficiency of transmission. The clock recovery circuit extracts the timing information from the data and synchronises the data. The final output of the system is the data and the clock, when necessary.

Now, let us discuss the different components of the link in a little more detail. The discussion would be mainly on the fibre and the electrooptic and optoelectric transducers as the rest of the link is standard electronic circuitry common to any other link.

### 1.2.1 The Optical Fibre

The optical fibre is made of glass - it transmits light waves through acting as the optical waveguide. The structure of the fibre is essentially simple - its consists of concentric layers of glass. The inner one is known as the core and the external one is known as the cladding. They can be differentiated in their optical nature - the cladding has a lower refractive index than the core.(Figure 1.2). The

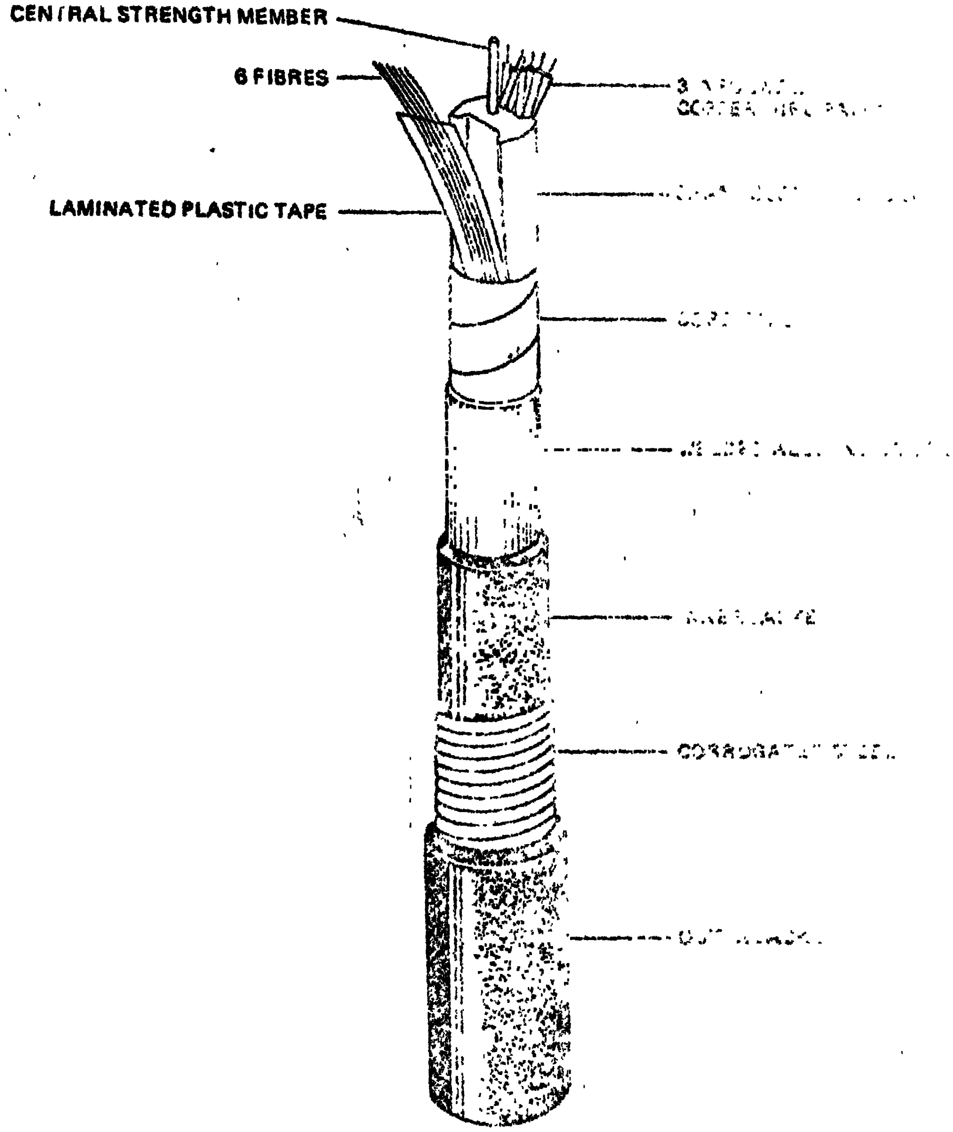



**Fig. 2 a)** – **Review of optical cables:** GTE cable. (After [5?])

FIG. 1.2

fibres are usually kept inside cables to protect them.

The nature of propagation of light waves in the fibre can be analysed with waveguide theory - but the basic properties can be explained from ray optics also. Consider the case shown in Figure 1.3. Consider a pencil beam of light incident on the cable - only three among the rays would elucidate the different possibilities. The rays undergo refraction at the entrance of the fibre - and also at the core and cladding boundary. As the core is optically denser than the core, the rays would be diffracted away from the normal. In case of ray (2), the angle of incidence is such that the critical angle is exceeded, it is total internally reflected back to the core. The ray (1) goes straight through the core as it is along the axis of the core. Ray (2) would undergo innumerable reflections at core-cladding boundary and proceed almost unattenuated. For ray (3), the incidence angle is less than the critical angle, and it is lost in the cladding - this type of rays constitutes the leaky mode of transmission and are essentially the cause of attenuation in fibres.

There is another feature of the fibre that tend to limit its performance like the attenuation. This is known as dispersion. It can be explained qualitatively from the above discussion. Consider rays (1) and (2) - both of them are propagating along the fibre. Though they might be

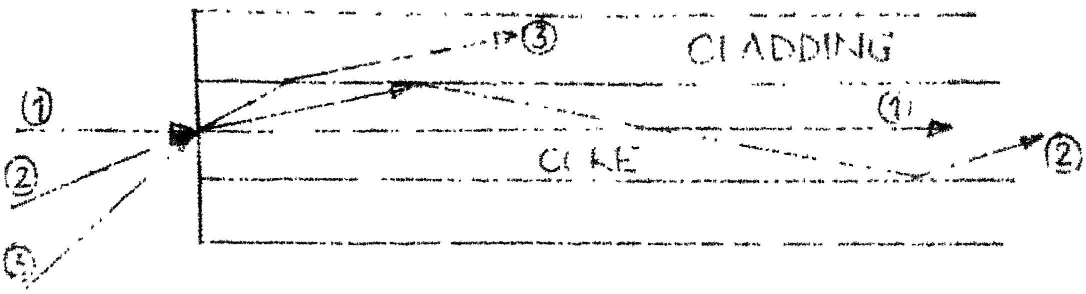



FIG. 1.3 TRANSMISSION OF LIGHT IN OPTICAL FIBRE

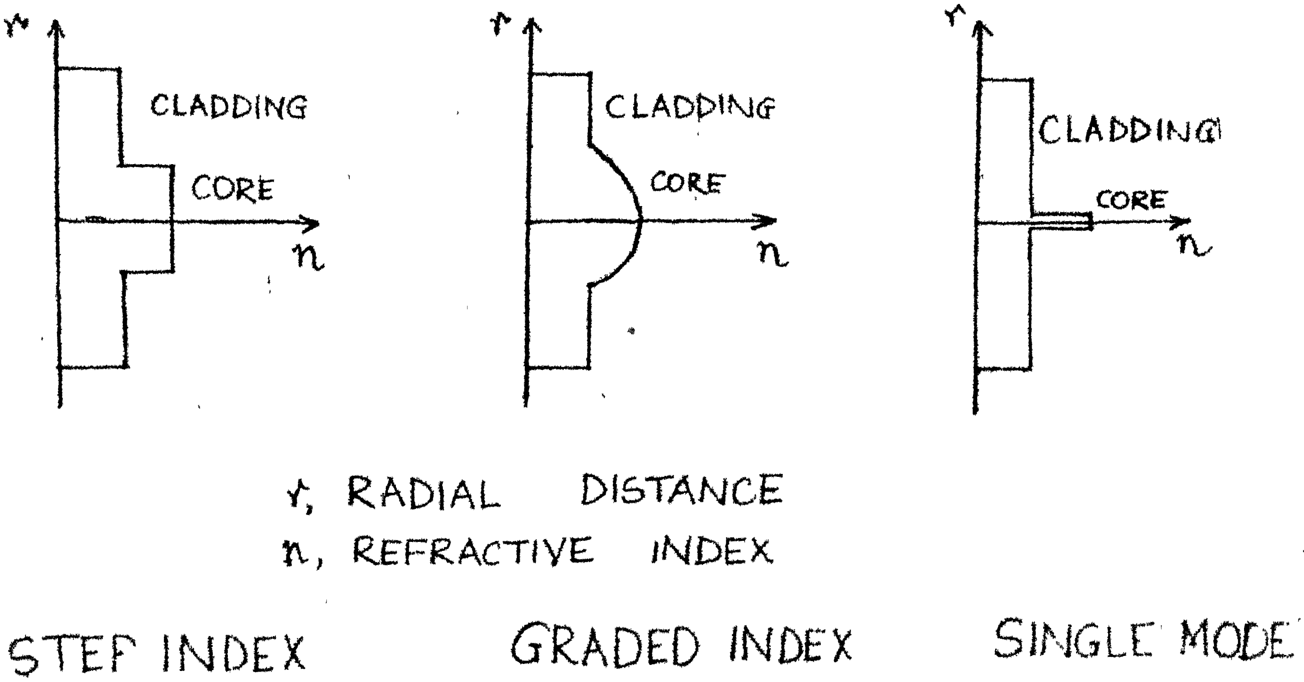



FIG. 1.4 DIFFERENT KINDS OF FIBRE

starting at the same instant at the input, they would arrive at different times due to the difference in distance travelled. Thus, an unit impulse in time would result in a broadened pulse at the output. This effect tends to limit the bandwidth capabilities of the fibre tremendously.

Modifications in the fibre structure have been proposed to counteract this affect. The fibre that has been considered here is called a step-index fibre - as the refractive index from the core to the cladding change in a step at the boundary. They are most affected by dispersion. Modifications of this type of fibres have resulted in two other kinds of fibres that reduce dispersion effects. One of them is called a Graded Index fibre, in which the transition of refractive index from the core to the cladding is smooth - this results in a focussing effect of the light waves travelling through the fibre. The rays, travelling straight or bent at the core-cladding boundary tend to move at different velocities. The effect of dispersion is minimised.

The other modification has resulted in single mode fibres — where the core is so small that only one ray - ideally - is allowed to pass through. The dispersion effect is thus almost totally absent. These fibres are best in high data rate applications - but their high cost and difficult handling, like alignment and coupling, have

restricted their use in other systems with moderate data rates. In the present work, graded index fibres have been used because they give very satisfactory performance at low cost and data rates upto several tens of Mbps.

### 1.2.2 Sources and Detectors

The sources and detectors are the transducers that are used to interface the electronic circuits with the optical systems. The sources and the detectors are thus essential parts of the link along with the fibres. The dimensions of these devices are such that they can be coupled with fibres with maximum efficiency.

The sources are mainly of two kinds - LED and Laser Diodes. LED's can be operated more easily than the Laser diodes - but they have lower power output. Laser diodes also have smaller spectral width than the LEDs and that reduces the dispersion effect in the optical fibres. This dispersion is known as Material Dispersion.

The detectors are also of two kinds — PIN and APD. APD has more electrical output at a given optical level than PIN detectors because of the avalanche multiplication in them. This makes them more noisy too - because the multiplication is a random process. APDs are more popular in applications requiring high sensitivity - whereas PIN detectors are used where low noise is more

important. The APD detectors need an extra stabilised bias voltage unlike PIN.

1.2.3 Connectors

Connectors are used to couple the fibre with the sources, detectors and also with other fibres. Because of the small sizeof the fibres, it would be impossible to align and fix them without these connectors.

1.3 <u>Data Rate Limitations in a Fibre Optic Link</u>

As discussed earlier, a fibre has two basic limitations — attenuation and dispersion. The attenuation in the fibre limits the repeater spacing in the link - the repeater spacing can be maximised with low loss fibres, high power sources and highly sensitive detectors. The effect of data rate on attenuation is meagre - except in the extreme case, where at very high data rates, the reciver sensitivity is decreased. Thus, the attenuation factor does not limit the data rates in the fibre for a given link.

The dispersion limits the bandwidth of the fibre mainly. There are basically two kinds of dispersion - the intermodal dispersion and the material dispersion. The intermodal dispersion occurs in multimode fibres and has been discussed earlier. The material dispersion is produced by the properties of the fibre and the impurities in it. It is also dependent on the nature of the source - being

directly proportional to its spectral width. Thus, a more coherent source like the Laser reduces the dispersion greatly.

The other factors that delimit the bandwidth are the rise time of the source and the detector. Typically, in case of LED sources, the rise time varies from 15 to 20 nsecs and with the Laser diode, 1 to 2 nsecs. In case of detectors, the rise time of APD and PINS are of the order of 1 to 2 nsecs.

Above all these, the finite bandwidth effect of the transmitter and the receiver would come into picture in the consideration of the data rate of the complete link. In this discussion, the optical link only is considered.

In the present experimental work, a LED source and a PIN detector were used A length of 2.45 km of graded index fibre was used in the experiment. For such a length, attenuation limitations does not arise.

The total rise time in the optical system is

$$T_R = 1.1\sqrt{t_{rs}^2 + t_{rd}^2 + t_{imod}^2 + t_{mat}^2}$$

where,

$t_{rs}$ : rise time of the source

$t_{rd}$ : rise time of the detector

$t_{imod}$ : intermodulation dispersion

$t_{mat}$ : material dispersion.

For the fibre used in the experiment, $t_{imod} = 1$ ns/km and typically, with LED sources, $t_{mat} = 3.5$ nsec/km.

Thus, in the present case, we have, $T_R = 23$ nsec.

For NRZ data, the rise time permitted would be a maximum of 70% of the clock period; hence, the minimum time period permissible is 32.85 nsec.

Thus, the maximum data rate possible is 30.4 Mbps.

A transmitter capable of operating at high data rate was available in the laboratory. The work was mainly to make a receiver that can utilize the channel capacity to the maximum.

## 1.4 Scope of Present Work

The present work essentially deals with the implementation of a low noise preamplifier that would follow the detector, because the amplifier specifies the main characteristics of the receiver such as sensitivity and data rates. The design and implementation of the amplifier is made critical by the fact that it should be low noise and high bandwidth at the same time.Several circuit configurations of the amplifier was tried out before arriving at the operational one.

A crystal controlled clock recovery circuit was also implemented to be used in conjunction with the preamplifier at the receiver. The crystal controlled clock shows

very small drift with time and temperature - and if it is coupled with a crystal at the transmitter also, the system would become very stable. This was implemented with the system.

## 1.5 Thesis Layout

The design consideration for the front end preamplifier that has led to the choice of a particular device and amplifier configuration is the topic for discussion for the next chapter. The design and implementation of the preamplifier circuit used has been discussed in the Third Chapter. The complete link that was used in the experiment and the performance comes in the Fourth Chapter. Finally, the conclusion and suggestions for further work appear in the concluding chapter of the thesis report.

# Chapter 2 Front End Design Considerations

## 2.1 Important Receiver Characteristics

Fibre optic systems are now becoming popular in a great number of communication systems like point to point links, long haul links and local area networks. Though the essential nature of the link does not change for different applications, the structure of the receiver changes drastically. The requirements for the receiver changes - sometimes in between conflicting features. Recently, there has been a drive to incorporate as many features as possible in a receiver such that it can handle a variety of links.

The most important criteria for fibre optic receivers are sensitivity, dynamic range and acquisition time [4,5] . There are other requirements like data format transparency, but they are not so important in the present discussion. They mainly come into consideration in computer networks and related fields.

The sensitivity of a receiver indicates the minimum signal level, with which it can operate within a safe limit of performance, for example, with BER less than $10^{-9}$. It is the most important criterian for long haul links such as submarine links where the spacing of the repeater has to be maximised. It is an important feature in other links also.

Acquisition time is an important criteria for point to point links — this specifies the time taken for an idle receiver, receiving no data, to stabilize itself on a incoming stream of data. The clock must also be able to synchronise with the transmission clock during this period. For bursty modes of data, fast acquisition time is a must.

In case of Local Computer Networks, fast acquisition time is an important criteria — dynamic range is also significant in this case. Dynamic range is the input levels within which the receiver can perform satisfactorily. In computer networks, where the nodes may be positioned randomly, a wide dynamic range is an essential requirement.

Each of these criteria result in the modifications in receiver design for some particular application. In our case, the link that is mainly of interest is the point to point link. Thus, the discussion would be mainly on receiver sensitivity and its variation with bit rate.

The front end amplifier and the detector are the most important constituents of the receiver on which the entire performance of the receiver depend. The rest of the amplifier, constituting mainly of amplifier, decision circuit and filter stages, if necessary, are standardised communication systems.

In the present work, as an PIN detector has been used, the performance of the detector is also known. This leaves the front end amplifier only to be optimised for maximum data rates.

## 2.2 Receiver Configurations

Before the discussion of the receiver sensitivity, the receiver configuration that would result in the best performance of the receiver would be discussed. There are basically two kinds of configurations for front end amplifiers - high input impedence and transimpedence amplifiers [4,6]. Both of them are essentially current to voltage converters because the detector output signal is current and a voltage signal is better for processing at the output. Their schematic diagram are given in Figure 2.1.

In the first case of high input impedence amplifiers, the amplifier mainly acts as voltage amplifier. A voltage signal is produced when the current signal flows through a high impedence at the input of the amplifier. The most important advantage of this circuit is that it offers the lowest noice level and hence the highest detection sensitivity. A high input resistance results in both high current to voltage transfer ratio and low noise. But a high resistance affects the frequency characteristics of the amplifier, because the input time constant becomes too large. As a result, the pulse shapes are integrated. Hence, an

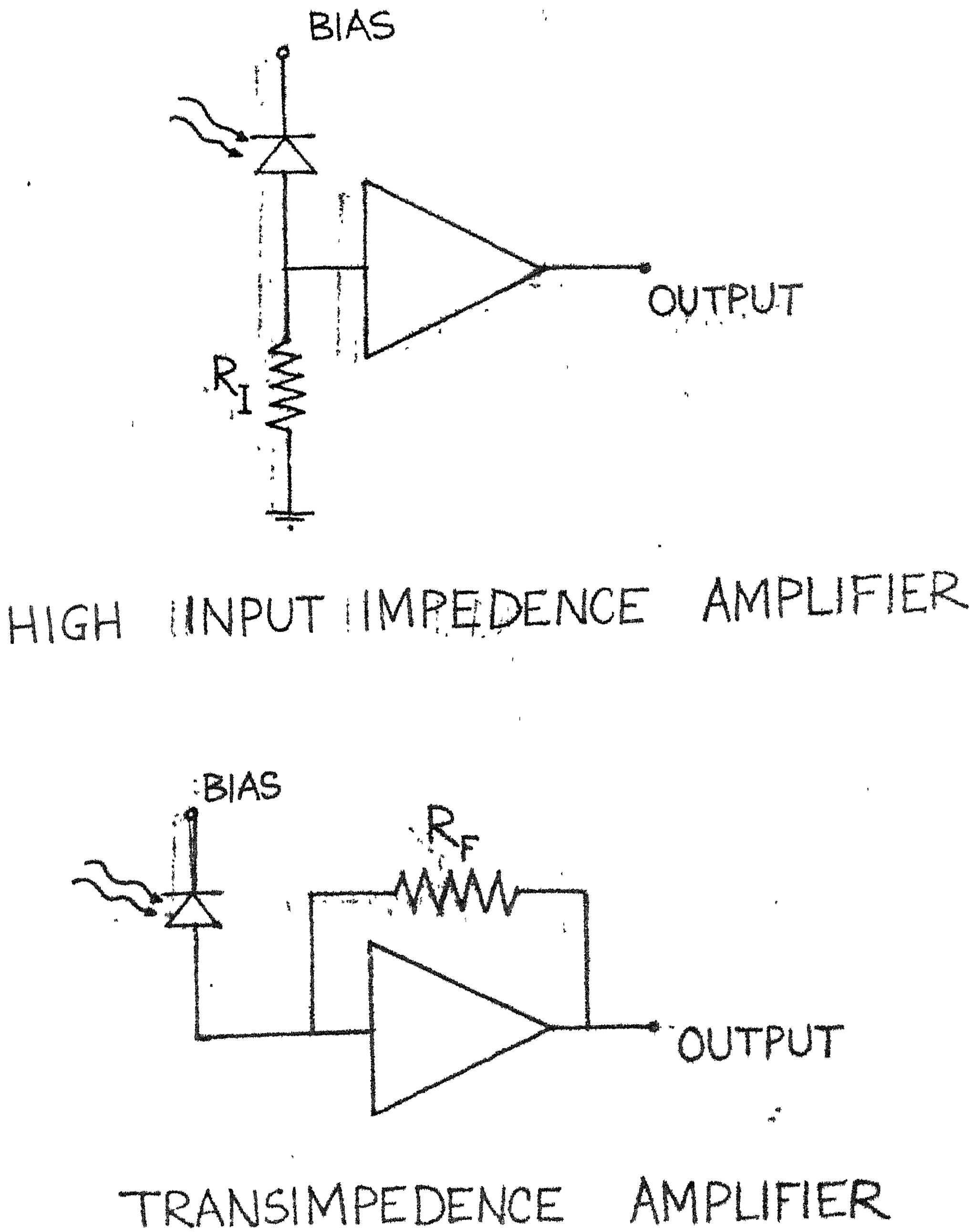



FIG. 2.1 PREAMPLIFIER CONFIGURATIONS

equilizer is used after the amplifier to restore the pulse shape. The equilization is often difficult — if the characteristics of the amplifier changes due to aging or replacement, the equilizer properties have to be changed in an adaptive manner to reduce the distortion.

The transimpedence amplifier does not suffer from this constraint because it is based on the concept of negative feedback. The amplifier bandwidth can be sufficiently increased to minimise distortion, and thus, to do away with the equilizer. In this case, the output voltage develops as the current flows through the feedback resistance. The disadvantage of the system is increased noise level at the input and decreased sensitivity. The comparatively smaller value of feedback resistance results in more noise than the high input impedence version. The dynamic range is more with transimpedence amplifier than high input impedence configuration.

In the present work, where an extremely low noise configuration was not very essential - the transimpedence configuration is used as the front end amplifier.

## 2.3 Choice of Front End Device

There are two types of devices that are used as Front End Amplifiers. They are Field Effect Transistors and Bipolar Function Transistors. The variation of sensitivity of the receiver with data rates would be discussed for both these devices - the analysis would help to clarify the choice

of device in the present experiment.

### 2.3.1 Receiver Sensitivity

As defined earlier, the sensitivity of the receiver is given by the minimum input power required for it to operate. Mathematically, it can be considered as the minimum average detected output power for a particular Bit Error Rate. This consideration assumes the quantum efficiency of the detector to be 100% and thus can be directly measured at the input of the receiver. The consideration would thus be based on the noise characteristics of the detector and receiver preamplifier only.

The noise equivalent power, in terms of equivalent input noise current power of the amplifier, $i_{na}$, and the dark current noise power, $i_{nd}$, is given by, [6 , 7],

$$<i_n^2> = <i_{nd}^2> + <i_{na}^2>.$$

The dark current noise power can be written as [8],

$$i_{nd}^2 = 2e\, I_{du}\, BI_2 + 2e\, I_{dm}\, G^2 F(G)\, BI_2$$

where,

B : Bit Rate

$I_2$ : Constant depending on the input and output pulse shapes

$I_{du}$ : Unmultiplied dark current

$I_{dm}$ : Multiplied dark current

G : Avalanche Gain

In case of PIN detector, G is unity. F(G) is also unity and $I_{dm} = 0$, so we need to consider the first term only.

It has been calculated [6], for a given error rate, the sensitivity, in terms of average detected optical power input, is given by

$$\bar{P} = Q \frac{h\eta}{e} [QeI_2B + \sqrt{<i_{na}^2>}]$$

where,

$h\eta$ : photon energy

e : electronic charge

Q : a parameter related to desired error rate, $P[E] = \frac{1}{\sqrt{2\pi}} \int_Q^{\infty} \exp(-x^2/2)dx$.

For a standard BER, $P[E] = 10^{-9}$, $Q = 6$.

For a PIN detector, where the thermal noise $<i_{na}^2>^{1/2}$ generally dominates over the shot noise factor, $QeI_2B$, the expression reduces simply to

$$\bar{P} = Q \frac{h\eta}{e} \sqrt{<i_{na}^2>} .$$

Thus, the sensitivity is solely dependent upon the noise at the input of the receiver amplifier.

### 2.3.2 FET Front End

For an FET front end, the input equivalent noise current power is given by, [6,7,9]

$$\langle i_{na}^{2} \rangle = \frac{4kT}{R_f} I_2 B + 2e\, I_L I_2 B + \frac{4kT}{g_m} (2\pi C_T)^3 f_c I_f B^2 + \frac{4kT}{g_m} (2\pi C_T)^2 I_3 B^3$$

where,

| | | |
|---|---|---|
| $k$ | : | Boltzmann's constant |
| $T$ | : | Absolute temperature |
| $R_f$ | : | Feedback resistance for transimpedence amplifier |
| $B$ | : | Operating Bit Rate |
| $I_L$ | : | Total leakage current (includes the gate current and the dark current of the photodetector) |
| $g_m$ | : | Transconductance |
| $C_T$ | : | Total input capacitance |
| $f_c$ | : | 1/f noise corner frequency of FET |
| | : | Device constant [1.03 for Si, 1.75 for GaAs] |
| $I_2, I_3, I_f$ | : | Constants which depend upon the input optical pulse shape to the receiver and the output pulse shape. |

For NRZ coding and Raised Cosine Output pulse, $I_2 = 0.562$, $I_3 = 0.068$, $I_f = 0.184$.

The first term is the contribution of thermal noise of the feedback resistance. The second and third terms are due to the shot noise and the 1/f noise. The last noise

term is caused by channel thermal noise and induced gate noise of the device.

The equivalent noise power can be rewritten as follows:

$$< i_{na}^2 > = \frac{4kT}{R_f} I_2 B + 2e\, I_L I_2 B + \frac{4k}{g_m} (2\pi C_T)^2 . (1 + \frac{I_f}{I_3} \cdot \frac{fc}{B}) I_3 B^3 .$$

For Si JFETS, the gate leakage current is negligible and the second term may be dropped. Also, the 1/f corner frequency is very small compared to the bit rate.

But, due to their low value of transconductance and high input capacitance the effect of the third term is quite high. Thus, though they are good at low frequencies, their performance degrades sharply at high frequencies, with increasing bit rate. For a typical value of $g_m$ as 5 mmh and a capacitance of 0.5 to 1.0 pf, they can hardly be pushed beyond 40 Mbps [5].

The analysis of GaAs MESFETs has shown that their properties are far better than Si JFETs and they can be operated upto data rates of 500 Mbps without any problem. However, such devices are not available here.

### 2.3.2 BJT Front End

For Bipolar Junction Transistors, the input equivalent noise current power is given by [5,6],

$$\langle i_{na}^2 \rangle = \frac{4kT}{R_f} \cdot I_2 B + 2e\, I_b I_2 B + \frac{2eIe}{g_m^2} (2\pi C_T)^2 I_3 B^3$$
$$+ 4kTr_{bb'} (2\pi C_{dsf})^2 I_3 B^3$$

where,

$I_b, I_c$ : Base and collector currents

$r_{bb'}$ : Base spreading resistance

$g_m$ : Transconductance, given by $I_c/V_T$ where $V_T$ is volt equivalent of temperature, $V_T = \frac{kT}{e}$.

e : Electronic charge

$C_{dsf}$ : Stray and Feedback Capacitance.

The total capacitance G is given by $G = C_d + C_{b'e} + C_{b'e} + C_f$, where

$C_d$ : detector capacitance.

$C_{b'e}$, $C_{b'e}$ : transistor junction capacitance in the hybrid-$\pi$ model.

The capacitance in BJT's is also a function of collector current - the capacitance changes as the base width changes due to collector current. $C_{b'e}$ can be subdivided into two parts as

$C_{b'e} = C_{je} + \alpha I_c$, where $C_{je}$ is the emitter junction space charge capacitance and $\alpha$ is the conversion constant from current to capacitance and is given by

$= \tau_F / V_T$ where $\tau_F$ is the forward transit time in the transistor.

The total capacitance can thus be written as

$C_T = C_o + \alpha I_c$ where $C_o$ is the zero bias capacitance.

In the expression for noise power, there are a number of terms varying directly or inversely as collector current and hence the noise power can be minimised for a particular $I_c$. The noise power can be differentiated with respect to $I_c$ and put to zero to calculate the optimum $I_c$. It comes out to be, [5]

$$I_{c\,opt} = 2\pi C_o V_T \, B \sqrt{\frac{I_3 \beta}{I_2}} \left[ \frac{1}{\sqrt{1+(2\pi\alpha V_T B)^2 \beta I_3/I_2}} \right]$$

where $\beta$ is the current gain of the transistor, $\beta = I_c/I_b$.

The noise power at optimum collector current is given by

$$i_n^2 = \frac{4kT}{R_f} I_2 B + 8\pi kT \, C_o \sqrt{\frac{I_2 I_3}{\beta}} \, B^2 + 4e\alpha C_o (2\pi V_T)^2 I_3 B^3 + 4kT \, r_{bb'} (2\pi C_{dsf})^2 B^3 .$$

In the transimpedence design, the feedback resistance $R_f$ would be varying inversely with Bit Rate to maintain maximum bandwidth. Hence, the first two terms would vary with square of the change in Bit Rate. It can be easily compared to FET in which the effective noise variation is proportional to the cube of the data rate.

For all these reasons, a BJT front end amplifier was decided upon as the receiver preamplifier in the implementation.

# Chapter 3 Receiver Amplifier Design and Implementation

## 3.1 Preamplifier Configuration

In the previous chapter, the choice of the device and the structure of the preamplifier have been discussed. The particular circuit that has been implemented would be discussed. But,before that, some other aspects of preamplifier configurations would also be dealt with.

Basically, there can be two kinds of amplifier stages — direct coupled or dc coupled for short and capacitance coupled or ac coupled [4]. The dc coupled amplifier can amplify any signal starting from zero frequency or dc to the maximum range of the amplifier. The acquisition time of the receiver is also small. But, the main problem lies in the bias stability. Also, the dynamic range may be affected because the input is biased for maximum sensitivity and a input signal overdrive might distort the output waveform.

Ac coupled amplifiers are, in this manner, better because there is no bias problems — there can be separate bias for each of the amplifier stages. The dynamic range is also better because the amplifier is always biased at the centre of the load line irrespective of the input signal. The main disadvantage of this type of amplifier is that they require a zero or a very low running average of the input signal. Otherwise, there is the problem of 'baseline wander',

that is, the reference input level changes with time — which may result in the error in detection when the noise margin is reduced.

Dc coupled stages with feedback combine the best of both the worlds. They feature feedback for minimisation of the bias stability problem and also the better low frequency performance of the dc amplifier. Thus, the dc amplifier involving feedback was implemented for preamplifier configuration. Though, when the preamplifier was coupled to a linear amplifier for more amplification, the coupling had to be capacitive because there was no way to stabilize the bias of the linear amplifier.

### 3.2 Choice of Preamplifier Device

After having chosen Bipolar Junction Transistor as the amplifier active component - the particular choice of the present experiment was made. It is obvious, at high frequencies, monolithic transistorsare the best performers. They have much less stray capacitive effects than the discrete components. Also, a number of them comesin a single chip, and it is called a transistor array. The transistors in the array have matched characteristics. Hence, a silicon microwave array was the preamplifier choice device.

The two monolithic arrays that were available were from National Semiconductor, LM 3046 and from RCA, CA3127E.

The cutoff frequency, which is the most important criteria for the choice of high frequency device, was 560 MHz for LM3046 and over 1 GHz for CA3127E. Hence, the circuit was implemented with RCA3127E.

### 3.3 Preamplifier

The preamplifier that was implemented can be broadly divided into two parts — a transimpedence amplifier and a voltage amplifier. Basically, the complete amplifier may be visualised as an transimpedence amplifier because the input signal is a current signal and the output is obviously voltage. But, in practice, the transimpedence amplifier is followed a voltage amplifier — for maximised bandwidth the transimpedence has been kept minimum, which has resulted in decreased output voltage. Thus, an amplifier stage was added to boost it up (Figure 3.1).

#### 3.3.1 Transimpedence Amplifier

A transimpedence amplifier has a configuration as shown in Figure 3.2.A. The output voltage is proportional to the product of the signal input current and the feedback resistance, $v_o = I_s R_f$. But, this is an ideal transimpedence amplifier where the amplifier A is considered to be ideal in all respects. It has infinite input impedence, zero output impedence and infinite gain at the frequency of operation. A practical amplifier with the nonideal effects has been shown

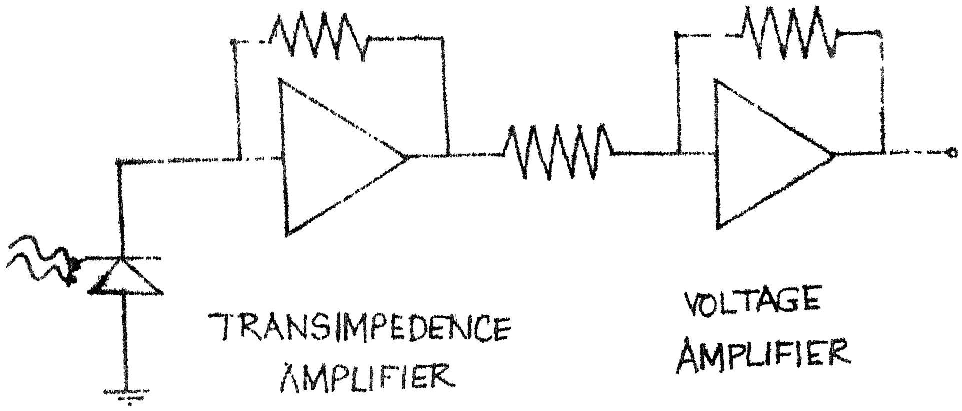



FIG. 3.1 PREAMPLIFIER BLOCK DIAGRAM

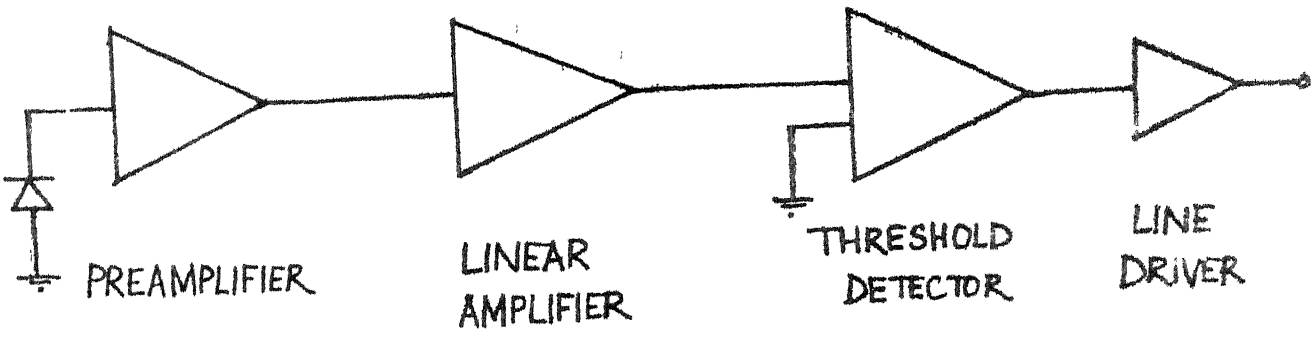



FIG. 3.4 RECEIVER AMPLIFIER BLOCKS

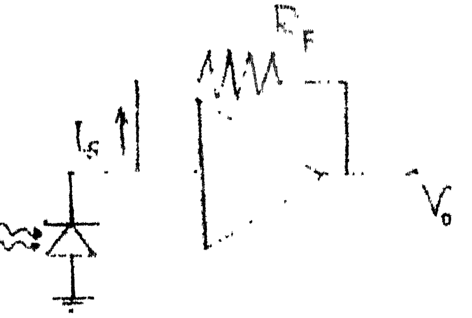



FIG. 3.2A

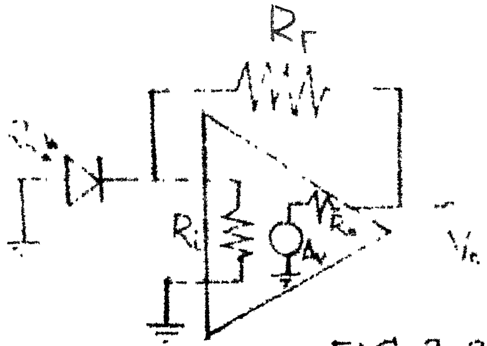



FIG 3.2B

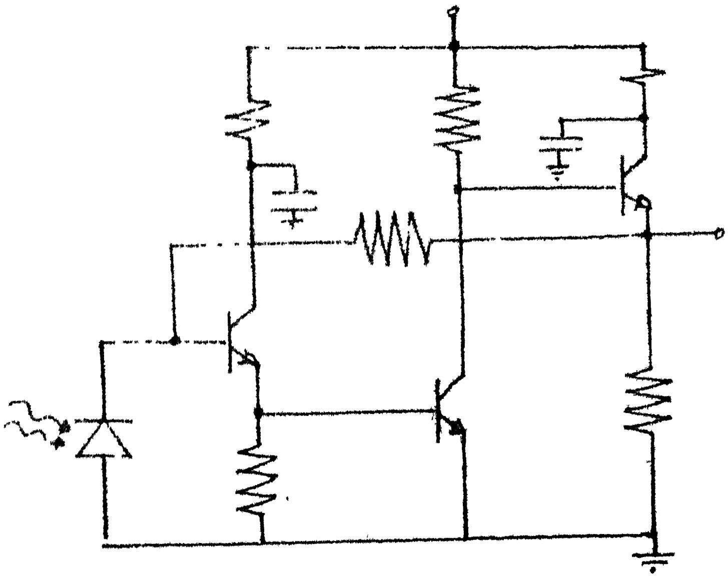


FIG. 3.2C

FIG. 3.2

TRANSIMPEDENCE AMPLIFIER

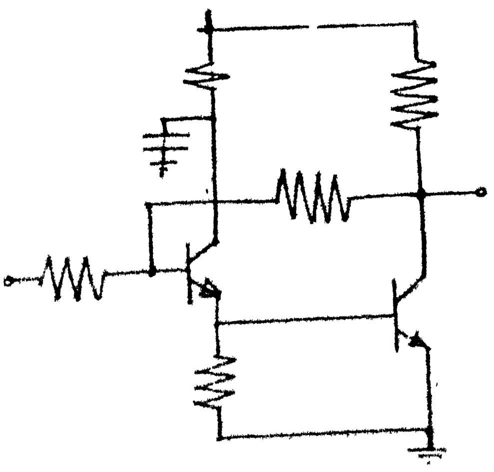


FIG. 3.3

VOLTAGE AMPLIFIER

in Figure 3.2.B. In the practical implementation, the conditions can be modified slightly to say, the input impedence should be much higher than the feedback resistance, the output impedence much smaller and the voltage gain is sufficiently high such that the gain is limited by the feedback resistance only.

A single transistor stage realisation of all these characteristics is almost impossible to achieve — hence, a three stage amplification has been done. The first stage the emitter follower, provides the buffering to the input and gain stages and also result in high input impedence. It has a superior noise property also which is discussed below. The second stage supplies the necessary gain - the last stages gives the output buffering and the feedback. As complete amplifier is dc coupled, the stability in bias also provided by feedback. This part is shown in Figure 3.2.C.

In the previous chapter, the noise properties of the preamplifier device has been discussed. It was noted that for BJT, at some collector current the noise power is a minimum - the expression for optimum collector current is

$$I_{opt} = 2\pi C_o V_T B \sqrt{\frac{I_3 \beta}{I_2}} \left[ \frac{1}{1+(2\pi\alpha V_T B)^2 \beta I_3/I_2} \right]^{1/2}$$

For low bit rates, $B \leqslant 100$ Mbps, the second term may be neglected and the expression rewritten as

$$I_{opt} = 2\, C_o V_T B \sqrt{\frac{I_2 \beta}{I_2}} \, .$$

With a typical $\beta$ of 50, and $C_o$ around 1.0 pf, the optimum collector current for B = 10 Mbps to 50 Mbps lies around 0.1 mA [5].

In the present implementations the input stage was designed in such a manner that the collector current was $I_{c_1} = \frac{V_{BE2}}{R_{E_1}}$ and $V_{BE_2} = 0.7$ V, $R_{E_1} = 4.7$ k ; this gives a collector current of 0.13 mA which is quite close to the optimum value. Thus, the input stage was designed for minimum input noise also.

### 3.3.2 Voltage Amplifier

A two stage amplifier, as shown in Figure 3.3, have also a buffer stage at the input followed by a gain stage. The feedback as in the previous case, provides both the gain and bias stability.

Thus, a two level of amplification has been achieved with all the five transistors in the array.

## 3.4 Linear Amplification and Threshold Detection

As the output of the preamplifier is not sufficient for the threshold detector to operate unambiguously, the signal was further amplified by a video amplifier. (Fig.3.4). The linear amplification provided by this stage, consisting of a 733C, is sufficient for the comparator. 733 was chosen because it is cheap and a large bandwidth. The comparator that has been used is NE529.NE529 is a fast comparator chip with typical propagation delay of 10 nsec. The shaped output is passed through a TTL line driver to provide output buffering. As mentioned earlier, the last parts of the receiver that was discussed just now are common to any other communication channel and have already been standardised.

The complete receiver circuit appears in Figure 3.4.

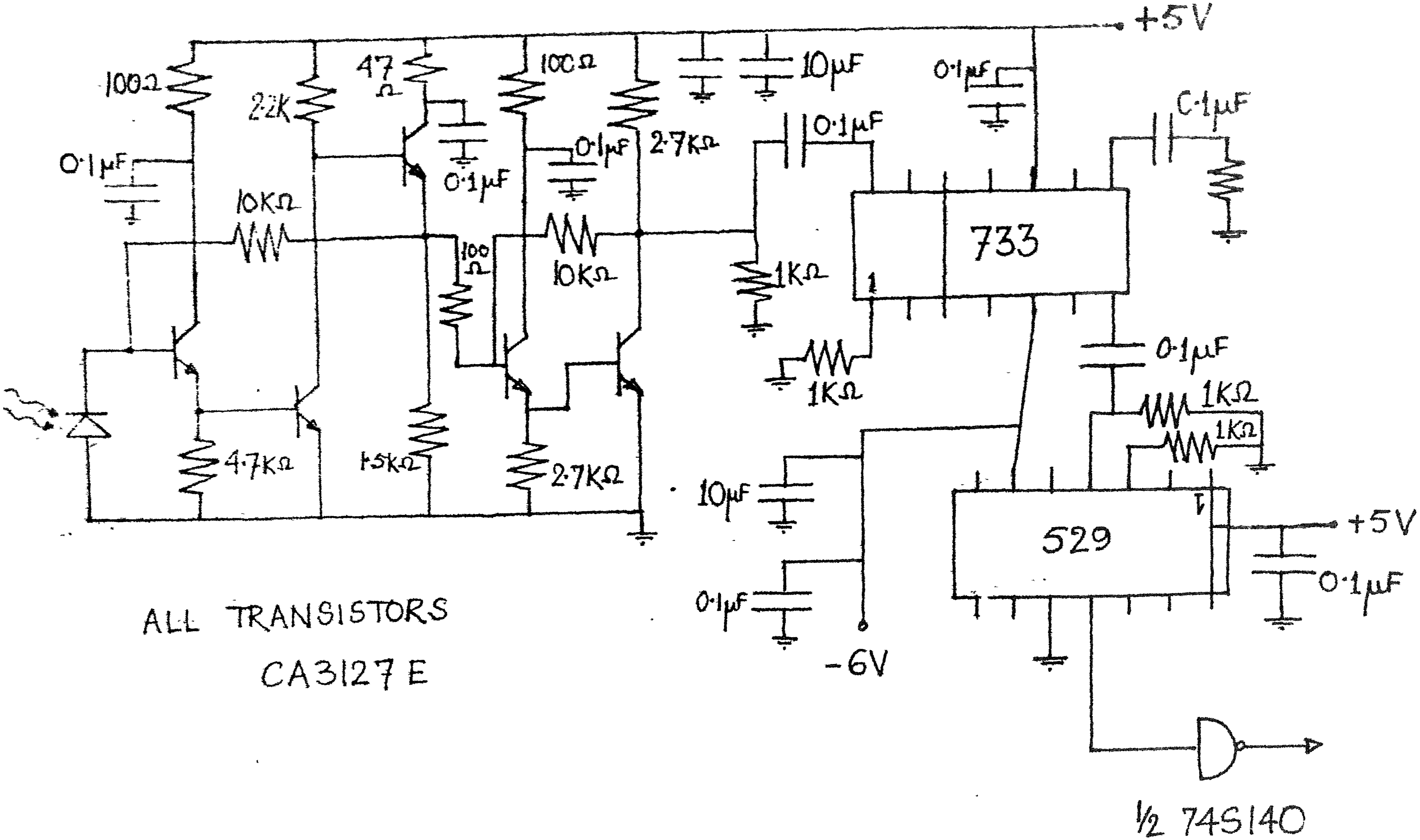



FIG 3.5 RECEIVER AMPLIFIER

# Chapter 4 : The Complete Link - Implementation & Performance

In this chapter, the complete link for which the receiver amplifier was fabricated would be discussed and the performance of the amplifier in the given link would be analysed. The link, as shown in Figure 4.1, is similar to that shown and discussed in Chapter 1. The individual building blocks that were used in the link would be detailed in the rest of the chapter. The transmitter modules would be treated first and then the receiver modules would be dealt with.

## 4.1 Transmitter Building Blocks

### 4.1.1 Clock Generator

A crystal was used to generate the clock in the transmission of data in the link - this was basically due to the reason of clock stability. The crystal ensures that the clock would not drift much with time or temperature unlike any other source of clock. The importance of this building block would be appreciated when the receiver clock recovery circuit is discussed - it can only be commented here that it improved the acquisition time of the receiver.

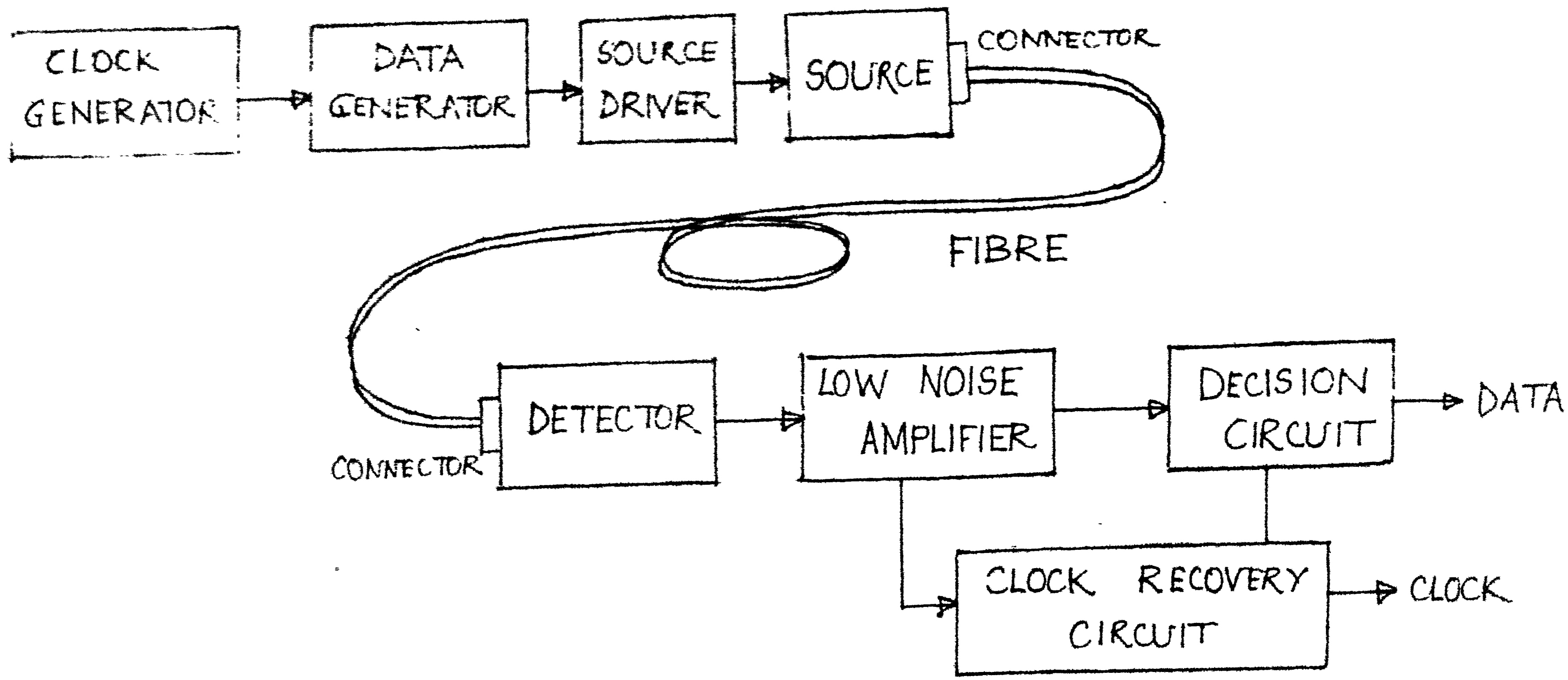



FIG. 4.1 EXPERIMENTAL FIBRE OPTIC LINK

4.1.2 Data Generator

As the link setup did not have a source of real-time data — a Pseudo Random Bit Sequence was used as the standard simulated data. A PRBS Generator was used as the source of data. It was a Hewlett Packard 3180 A, with the external crystal determining the clock. The data generator is a versatile one, with the capability of generating PN sequence upto a length of $2^{15} - 1$ bits. It can also transmit a 1010..... sequence, add any number of Zeros upto 100 in between the sequences, add errors etc.

4.1.3 Source Driver and LED

The source driver is the penultimate electronic device before the optical domain takes over in the link. The driver is used to supply current to the optical source and switch it on and off according to the input data. In the present setup for the link - an IIT module, type TXD 005C, containing the source driver and the LED has be used. This was used because it can deliver high power output — about -6dBm of optical power - this is achieved by switching about 300 mA of current in the LED source.

4.2 <u>Fibres and Connectors</u>

4.2.1 The Fibre

The optical fibre used in the experimental setup was a 2450m long fibre - this is also a product of IIT. The

dimensions of the core is 54 μm and of the cladding 126 μm. The attenuation in the fibre is 3.36 dB/km. and the dispersion is 1 ns/km. The fibre is a graded index fibre - as it was discussed in Chapter 1, this type of fibre optimises the performance and cost in the present operation.

4.2.2 The Connectors

The connectors that were used to couple the fibres with the source and the detector were numbered as OFP-101. The source, the detector, as detailed below and the connectors are all from ITT.

Due to the mismatch in the numerical aperture of the source and fibre connector at the input -the source aperture being much more than that of the fibre - only a fraction of the source output power could be coupled to the link. Typical low at the source connector was about 10 dB. At the receiver, on the other hand, there was almost no loss in coupling because the receiver aperture is much more than that of the fibre.

## 4.3 Receiver Building Blocks

4.3.1 The Detector

The PIN detector device number is OFDP-04. Among the important properties of the detector are its maximum power handling capability, the minimum power for

reliable operation, or sensitivity and the dark current. The sensitivity of the detector is -45 dBm and the maximum power it can handle safely is -20 dBm. The dark current is less than 0.180 nA - the effect of dark current on the performance of the detector was not high because at the power level required for the operation of the receiver, The dark current was about 1% of the signal output current. Due to the absence of avalanche gain as in APD, the sensitivity of the detector and the noise contribution was less than the former device.

### 4.3.2 Receiver Amplifier

The module has been discussed in detail in Chapter 3.

### 4.3.3 Clock Recovery Circuit

In digital telecommunication systems, the data that is transmitted through any channel has to be synchronised with a clock. The digital bits do not have any meaning if the clock information is not present. Thus, at the receiver of the digital transmission system, the clock must also reside with the data.

Usually, the transmission rate is known at the receiver also - there is a free running clock at the receiver which has to be synchronised with the incoming data rate. The output of the receiver might contain both the data and

the clock for some applications, or the data only which is changing at precisely known time intervals at the receiver - the clock information is implicit.

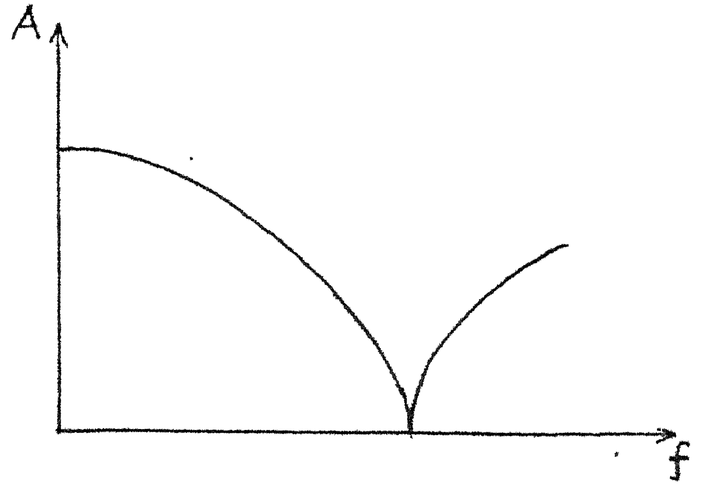
The synchronisation of the receiver clock with the transmitted clock has remained one of the most interesting problems in digital communication. A number of schemes have been proposed to achieve this purpose. The most straightforward solution is to send the clock along with the data. This disadvantage of this scheme is that it would result in wastage of bandwidth and power - but the advantage is that the receiver design would be quite simple.

The synchronising information might be sent in the form of conspicuous pulses at the end of some sequences - the receiver can use them to bring its clock in unision with the data. This scheme lies in between of the performances of the first and the next scheme - which is to obtain the clock from the data itself at the receiver. This would introduce complication in the hardware of the receiver structure - but the transmission capacity is used to the maximum. In high data rate systems, the last scheme is preferred. Sometimes, channel coding is introduced to fascilitate clock recovery at the receiver without wasting much of the bandwidth. In the present experiment, a data based clock recovery scheme is used but there is no channel coding.

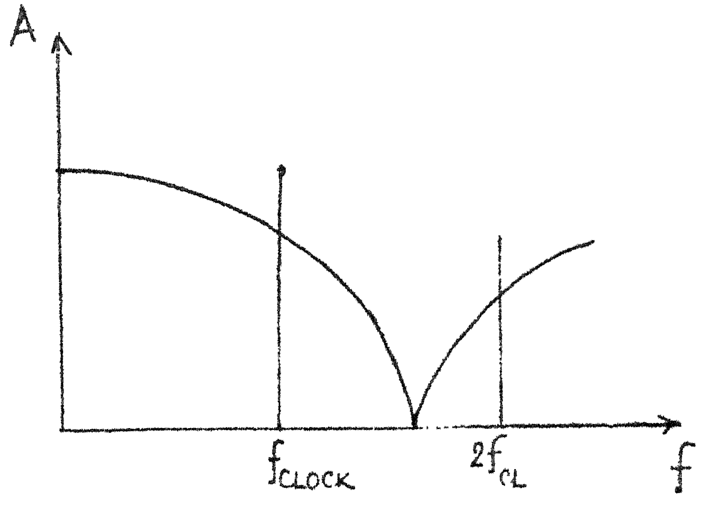
The transmission of data in the present case is in the form of NRZ coding for maximum utilization of bandwidth. This coding scheme results in a data stream that contains least information about the data rate. RZ coding incorporates more information about the clock but the effective data rate is only half of the NRZ stream for the same bandwidth. At the receiver, the NRZ data is converted back to RZ form. A number of ways are there to effect this conversion- the scheme that has been used uses only digital gate delays and an Ex OR [13]. The spectrum of the NRZ and RZ data also shows the difference of the two coding schemes, as shown in Figure 4.2. The spectrum of NRZ data is continuous - but for RZ, there are discrete components at the clock frequency and its multiples. A tracking filter is used to extract this single frequency from the 'self-noise' background of the continuous spectrum.

The Phased Locked Loop in Figure 4.3 is used as the tracking filter in this application - the individual blocks of the PLL are also shown. The free running frequency of the VCO is adjusted to be close to the transmission clock. The phase comparator compares the RZ data and the VCO output and the error in phase is fed back to the VCO for self correcting action. The active goes on until the clock and the RZ data are locked both in frequency and phase - the output of the VCO gives the clock.

FREQUENCY SPECTRUM




NRZ CODING




RZ CODING

FIG. 4.2 DIFFERENT CODING SCHEMES

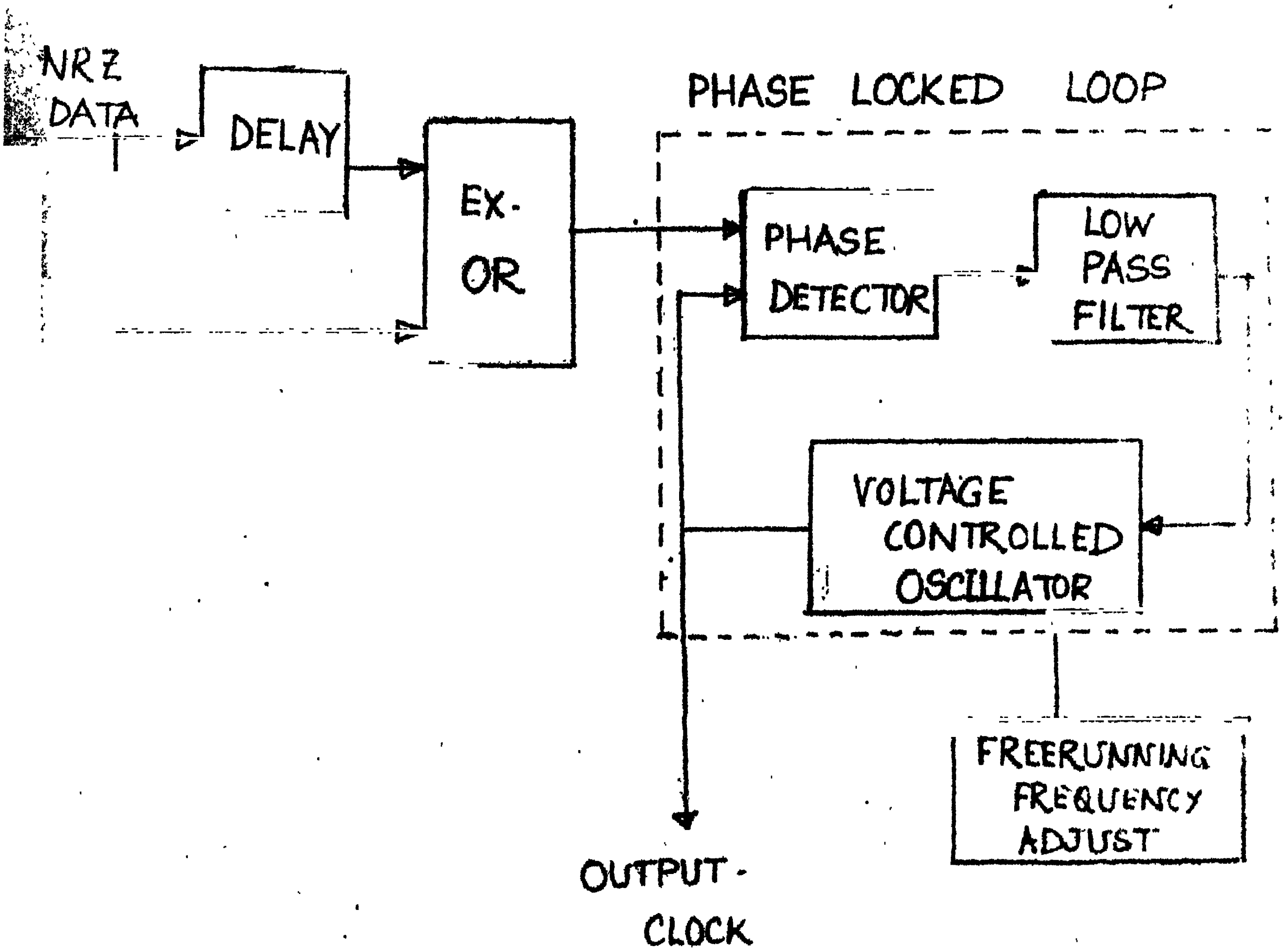



FIG. 4.3 CLOCK RECOVERY CIRCUIT

The low pass filter is used to minimise the bandwidth of the closed loop system. As the input signal has only one desirable component in the background of a continuous spectrum, the reduction of bandwidth removes the noise contribution of the other part of the spectrum except a small band around the frequency of operation.

The clock recovery circuit that was implemented for the experiment is given in Figure 4.4. The PLL used was XR215, which has a maximum free running frequency of 35 MHz.

The crystal clock input has resulted in a well stabilized clock - the clock with which the PLL can lock easily. The free running frequency of the PLL was also determined by a similar crystal as in the transmitter - this was also very stable. At any time, the synchronisation of the two clocks thus meant matching their phases only - as this can be achieved more quickly than frequency matching, the clock recovery time was very small.

## 4.4 Performance of the Link

### 4.4.1 Data Rate

The receiver preamplifier was tested for maximum frequency of operation — it was observed that the receiver does not perform reliably beyonds 20 MBps of data rate. Thus the frequency of operation can be maximum of 20 Mbps - but

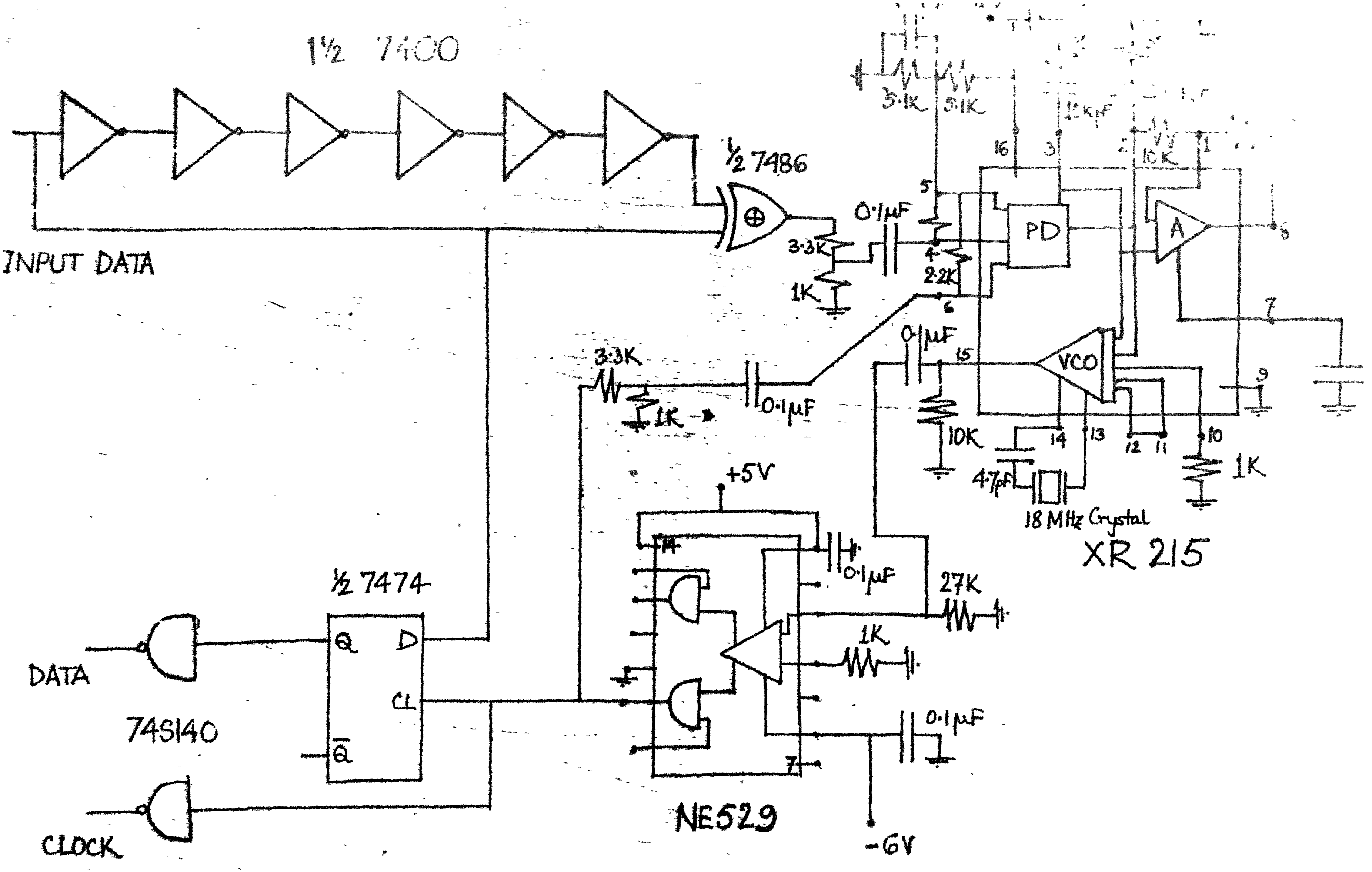



FIG. 4.4 BIT SYNCHRONIZER

crystals are not available at 20 Mbps. As crystals were available at 18 MHz, the link was setup at 18 Mbps using crystals at the transmitter and the receiver.

### 4.4.2 Error Performance at the Frequency of Operation

The following chart gives the variation of rate of error with input power at 18 Mbps NRZ data in the amplifier circuit.

| Power | Bit Error Rate |
|---|---|
| -26.9 dBm | No error during the period of observation (10 mts) |
| -27.7 dBm | Same as above |
| -28.4 dBm | Same as above |
| -29.2 dBm | Same as above |
| -30.7 dBm | $1.2 \times 10^{-7}$ |
| -31.7 dBm | $1.5 \times 10^{-6}$ |
| -32.2 dBm | $5.0 \times 10^{-5}$ |
| -32.7 dBm | $1.5 \times 10^{-3}$ |

The sensitivity of the receiver is around - 30 dBm. The dynamic range extends from - 20 dBm to - 30 dBm, the upper range is limited by the detector.

### 4.4.3 Clock Recovery Circuit

The clock recovery circuit was observed to be very stable with the crystal controlling its frequency. At 18 MHz, the pull-in range was approximately 10 KHz and lock range was observed to be around 25 KHz.

# Chapter 5 : Conclusion

A fibre optics receiver was fabricated in the present experimental work. Both the subsystems - the receiver amplifier and the bit synchroniser capable of operating at that frequency was implemented. The reasons for the choice of the particular devices and configuration of the two circuits and the performances obtained would be discussed here.

## 5.1 Receiver Amplifier

The preamplifier block was fabricated with a PIN detector and a BJT amplifier in the transimpedence configuratio. A linear amplifier and shaping circuitry complete the amplifier. The last two parts are usually not the limiting factor for the performance of the receiver - it is the preamplifier which is the vital part. The PIN detector was used with the preamplifier because of its low noise properties and also, because it was only device available at the time of the present work. This is also not the limiting factor as far as the sensitivity and bit rate is concerned.

The most important part of the implementation was the preamplifier implementation. As discussed in Chapter 2, the BJT front-end with the preamplifier was preferred because of its relatively better variation of

noise characteristics with bit rate. The noise characteristics of the receiver was not actually measured, the theoretical prediction has not been actually verified in practice. But, as the front end was able to function well with the optical system, an indirect inference can be drawn that the amplifier was really low noise. The receiver sensitivity analysis also confirms this.

The transimpedence configuration was used basically to remove the equilizer from the receiver. It would have been a must with high input impedence amplifiers. The transimpedence amplifier performed satisfactorily with the device used and at the bit rate of operation. In this case, the theoretical noise characteristics assume that the front end stage is connected in transimpedence mode and provides gain. Actually, a number of stages provide the gain and the feedback. The noise properties vary slightly from the predicted performance but the basic considerations remain the same.

The preamplifier in the receiver was optimised for maximum data rate - for better utilization of the capacity of the fibre. In a transimpedance amplifier, the increase in bandwidth can be achieved only with decreasing feedback resistance. The gain of the also decreased. The sensitivity of the receiver was affected by this.

Actually, the receiver amplifier has to be high bandwidth, high gain and low noise. The bandwidth, noise requirements has been discussed earlier. High gain is required because the detector output current ranges in fractions of microamperes and the output voltage of preamplifier should in the order of several milivolts for the following stage to operate well. If the device is of very high frequency nature - sufficient gain and bandwidth can be realised at the same time. In the present case, the device was not able to provide much gain over the bit rates of interest, hence the gain had to be decreased. The sensitivity was similarly decreased.

The sensitivity of the receiver was found to be -30 dBm. Though it was not as good as of many other fibre optic receivers - it would be still be able to maintain a standard 5 km length link with this receiver. The standard power output from the LED source is -5 to -6 dBm and after coupling losses are taken into account, the power coupled to the fibre is around -15 dBm. If the fibre attenuation is in the order of 2 to 2.5 dBm/km - this type of fibre is fairly common - the 5 km link would be able to function properly. The effect of dispersion is not considered here as a limitation of the repeater distance - the bit rate should be adjusted in such a case.

The dynamic range of the receiver was observed to be 10 dB. Here, it is limited by the detector maximum

power input and the sensitivity of the receiver. The detector might be replaced by another one that can accept more power. In cases like that, Automatic Gain Control (AGC) techniques might also be used to improve the dynamic range.

## 5.2 Bit Synchroniser

A crystal controlled bit synchroniser was implemented basically because of its rugged nature - the receiver clock can be synchronised with the transmitter clock even if the transmission is intermittant. There need not be a preample sequence before the data to synchronise the clock.

The bit synchroniser or the clock recovery circuit that was implemented corresponded exactly to the predicted performance. The system came into lock quickly even if the transmitter was not operating for a long period of time. The crystal resulted in a very sharply peaked characteristic of the voltage controlled oscillator and the filter - reducing the noise to a considerable extent. There are other advamtages of this configuration in the practical field also.

In usual case, the free running frequency of the VCO is determined by a capacitor. At high frequencies, the tolerance in the capacitance affects the circuit greatly. A

small variation in the value, for example, due to replacement, might drastically change the frequency and detune the receiver clock. The frequency might drift with change in characteristics of the VCO with temperature. Also more importantly, the supply voltage variation results in considerable variation in VCO frequency - a very stabilized power supply is required for the PLL.

With the crystal determining the free running frequency, all these effects are longer present. The crystal ensures that the frequency drift would be very little with time or temperature. The frequency becomes also fairly insensitive to the variation of power supply. Thus, the crystal resulted in a steadier performance of the VCO.

## 5.3 Suggestions for Further Work

The above discussion shows that the performance of the preamplifier was mainly limited by the device characteristics. The operations might be improved and higher data rate achieved with some modifications of the receiver structure and choice for a better device. The suggestions would be along the following lines:

1. A transistor array of higher cutoff frequency might be used as the front end device. A higher cutoff frequency, as discussed earlier, would be able to provide high gain and bandwidth and might drastically improve the performance.

2. A PINFET combination might be used as the detector. This device utilizes the low noise of PIN and low leakage properties of the FET to give superior performance at high frequencies.
3. A high input impedence and equilizer combination might also be tried out with suitable devices.
4. The PIN detector might be replaced by an APD. This would increase the sensitivity of the receiver by a factor of ten or more. At very high frequencies, the rise time of PINs that are available commercially might be limiting the performance of the optical link - an APD is a most at those applications.
5. Finally, the resistors and capacitors used in the amplifier might be replaced by better components suitable for high frequencies. Carbon resistors can be replaced by metal film resistances and ceramic capacitors by small tantulum capacitors.

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